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Direct-Conversion-To-Baseband for a Nonlinear FM Signal

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13. ABSTRACT *(Maximum 200 words)*

In this report, we examine both nonlinear FM signals and a "direct-conversion-to-baseband" process in a radar application system. We generate a phase predistortion function which modifies the transmitted linear FM signal and synthesizes the desirable effect of a sidelobe weighting function.

To avoid a mismatch in obtaining the in-phase and quadrature signal components at two separate processing channels, we directly convert a low IF signal to its baseband by a combination of sampling, filtering and decimation. A band-limited nonlinear FM signal with a low pulse compression ratio is considered and processed. The compressed signal is also obtained through a pulse compression filter matched to the transmitted signal.

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DIRECT-CONVERSION-TO-BASEBAND FOR A NONLINEAR FM SIGNAL

1. INTRODUCTION

For radar waveforms, such as a linear FM of relatively large time-bandwidth product, one can amplitude-weight the envelope of the transmitted signal or the receiver matched-filter frequency response to obtain a desired range sidelobe structure. It is practical and more efficient to apply weights at the receiver than at the transmitter, since control of the time envelope of a high power transmitter output is not feasible in most cases. An output signal-to-noise loss also occurs in time weighting cases as the average energy in the transmitted signal can not be adjusted to equal that of the unweighted signal. This is particularly true when the transmitter is peak-power limited.

One alternative in solving the above problem is to have some sort of phase predistortion that can correct amplitude or phase distortion in the receiver. For a nonlinear FM, an appropriate predistortion function is obtained which modifies the transmitted linear FM signal and synthesizes the desirable effect of a sidelobe weighting function.

In this report, we generate the spectrum-weighted nonlinear FM signals in which square roots of Taylor weightings are used. The waveforms are applied to a radar system where low pulse compression ratios are considered. We directly convert the received signal to its baseband through a simple process of sampling, filtering and decimation. The matched filter output results are also shown by using the reconstructed signal matched to the transmitted signal.

Figure 1 is a simple block diagram showing the pulse compression in radars. A narrow band signal is stretched out through an expansion network. The Doppler shifted radar return is received and processed with the receiver characteristics matched in phase to the transmitted signal for zero Doppler. The receiver spectrum is generally weighted in magnitude to reduce the output time sidelobes. Here we closely follow Fig. 2 to process the signal. We first generate a nonlinear FM signal with a carrier at 30 MHz. The phase characteristics of this waveform, as will be described in the next Section, assure that the time envelope of the signal is rectangular and the spectrum is Taylor shaped.

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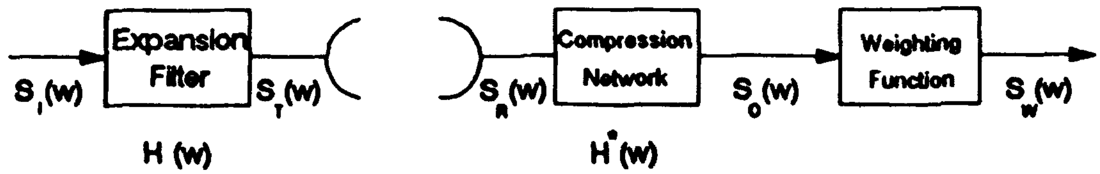

Fig. 1 The Radar pulse compression

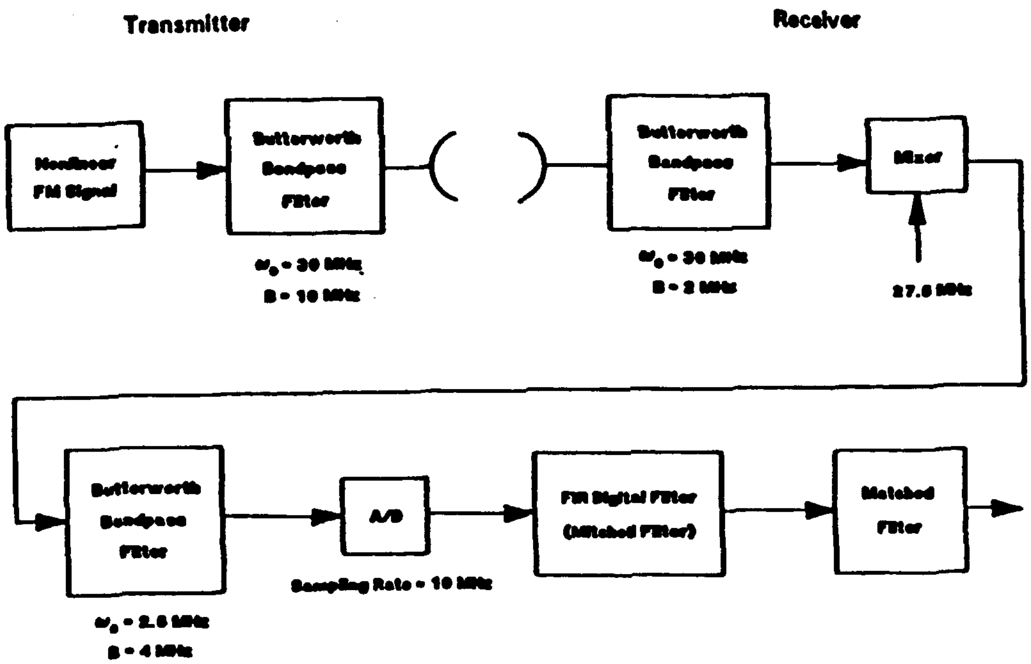

Fig. 2 A radar signal processing block diagram

The signal is transmitted after passing through a bandpass filter of 10 MHz.

In the receiver, the signal is band-limited by a 2 MHz Butterworth filter. To avoid a mismatch in processing signal in-phase and quadrature components in two separate channels, we down-convert the RF signal (actually a high IF in the simulations) to a low IF of 2.5 MHz. The signal is filtered by a 4 MHz Butterworth filter centered at 2.5 MHz and then sampled and digitized by an A/D converter with a sampling rate of 10 MHz. The real sample outputs are used to generate the complex components of the signal through a digital filter. The pulse-compressed signal is obtained through a matched filter with the filter response matched to the signal.

2. NONLINEAR FM SIGNAL

It was shown in Ref. 1 that appropriate phase predistortion can be employed to synthesize the desirable effect of a sidelobe weighting function as conventionally defined in Fig. 1. For a signal of f(t) (or, in frequency, $|F(\omega)| \exp(j\psi(\omega))$) such that its time envelope is rectangular and the spectrum is shaped as desired, the condition given by $|F(\omega)|^2 = k d^2\psi/d\omega^2$ must be satisfied. As a result, the waveform group delay becomes

$$t = -\frac{d\psi}{d\omega} = -\frac{1}{k}\int_0^{\omega} |F(\omega)|^2 d\omega . \tag{1}$$

As suggested in [1], we consider the square root of the Taylor weighting for $F(\omega)$, that is,

$$F(\omega) = (1 + 2\sum_m F_m \cos(m\omega/W))^{1/2} . \tag{2}$$

Here $\omega = 2\pi f$ is the angular frequency, W is the signal bandwidth, and the F_m's are the Taylor coefficients which can be calculated or obtained from a reference Table [2] for desirable sidelobe levels.

From (1) and (2), we can easily derive the group delay t as a function of frequency f. To explicitly describe a nonlinear FM signal, we present f in terms of t through a Fourier series presentation, that is,

$$f = f_o + W\frac{t}{T} + W\sum_n c_n \sin\left(2\pi n\frac{t}{T}\right). \tag{3}$$

Here W is the swept frequency during the time interval T, and the parameter W/T is conventionally defined as a constant k. The first 20 c_n for different Taylor weights: (a) -40 dB and $\bar{n}$=6, (b) -45 db and $\bar{n}$=10, and (c) -50 dB and $\bar{n}$=10 are derived and shown in Table 1. The plots of f/W vs t/T for the above cases (a) and (c) are shown in Fig. 3. Note that a special case of (3) can be obtained by setting all c_n's to zeros. This represents a linear chirp signal, i.e., $f(t)=f_0+kt$. The published c_n's such as in [1] or [2] are slightly different from we derived here where 200-point, instead of 24-point, harmonic analysis is applied. Only those C_n for -40 dB Taylor weights were shown in [1] and [2].

Denote by ρ, the pulse compression ratio. Then $\rho=T/(1/W)=kT^2$ with k=W/T. With (3) and the expression $\psi=\int 2\pi f dt$, the phase of a nonlinear FM is simply:

$$\psi = 2\pi \hat{f}_o \rho \hat{t} + \pi\rho\hat{t}^2 - \rho\sum_n \frac{c_n}{n}\cos(2\pi n\hat{t}) \tag{4}$$

where $\hat{t}$ and $\hat{f}_o$ are the time and the carrier frequency normalized to the uncompressed pulsewidth T and the bandwidth W, respectively.

The nonlinear FM signals (represented by cos $\psi(t)$) and their corresponding spectra when ρ=50 and ρ=5 are shown in Figs. 4 and 5, respectively. Here the C_n's for -50 dB and $\bar{n}$=10 are used. The signals shown in Figs. 4a and 5a consist of 256 samples and the normalized time extends from -1 to 1. Figure 6 shows the corresponding ambiguity diagram derived from Fig. 5. for ρ=5. The ambiguity diagrams of two other cases ρ=50 and ρ=13 are shown in Figs. 7 and 8, respectively. It can be seen that the range time sidelobe levels are significantly reduced with the increasing pulse compression ratios.

Table 1 -- Coefficients of Nonlinear FM signals

Nonlinear FM Coefficients	Cases		
	-40 dB, $\bar{n}=6$	-45 dB, $\bar{n}=10$	-50 dB, $\bar{n}=10$
c_0	-0.11419000E+00	-0.12388091E+00	-0.13296662E+00
c_1	0.39604966E-01	0.44902950E-01	0.49940077E-01
c_2	-0.20485607E-01	-0.24099349E-01	-0.27573738E-01
c_3	0.12532192E-01	0.15250478E-01	0.17890993E-01
c_4	-0.84144289E-02	-0.10575263E-01	-0.12696518E-01
c_5	0.59926926E-02	0.77751390E-02	0.95410396E-02
c_6	-0.44449255E-02	-0.59516317E-02	-0.74604202E-02
c_7	0.33963067E-02	0.46935917E-02	0.60067805E-02
c_8	-0.26545812E-02	-0.37873026E-02	-0.49463705E-02
c_9	0.21122884E-02	0.31118126E-02	0.41464046E-02
c_{10}	-0.17052770E-02	-0.25942659E-02	-0.35264694E-02
c_{11}	0.13932441E-02	0.21887064E-02	0.30353535E-02
c_{12}	-0.11498263E-02	-0.18649431E-02	-0.26390651E-02
c_{13}	0.95708821E-03	0.16023919E-02	0.23142783E-02
c_{14}	-0.80256353E-03	-0.13866427E-02	-0.20445227E-02
c_{15}	0.67735136E-03	0.12073694E-02	0.18178612E-02
c_{16}	-0.57488246E-03	-0.10569483E-02	-0.16254831E-02
c_{17}	0.49039477E-03	0.92966653E-03	0.14607797E-02
c_{18}	-0.42023009E-03	-0.82116433E-03	-0.13186573E-02
c_{19}	0.36153032E-03	0.72805533E-03	0.11951163E-02

3. FILTERING, SAMPLING, AND DIGITIZING FOR THE NONLINEAR FM SIGNAL

Before the complex nonlinear FM can be reconstructed through a simple finite impulse response (FIR) filter, the received radar signal is practically band-limited and reshaped at different processing stages (see Fig. 2). Here we show the resulting waveforms as affected by various filtering and sampling processes. In Section 4, the down-converted and digitized signal at

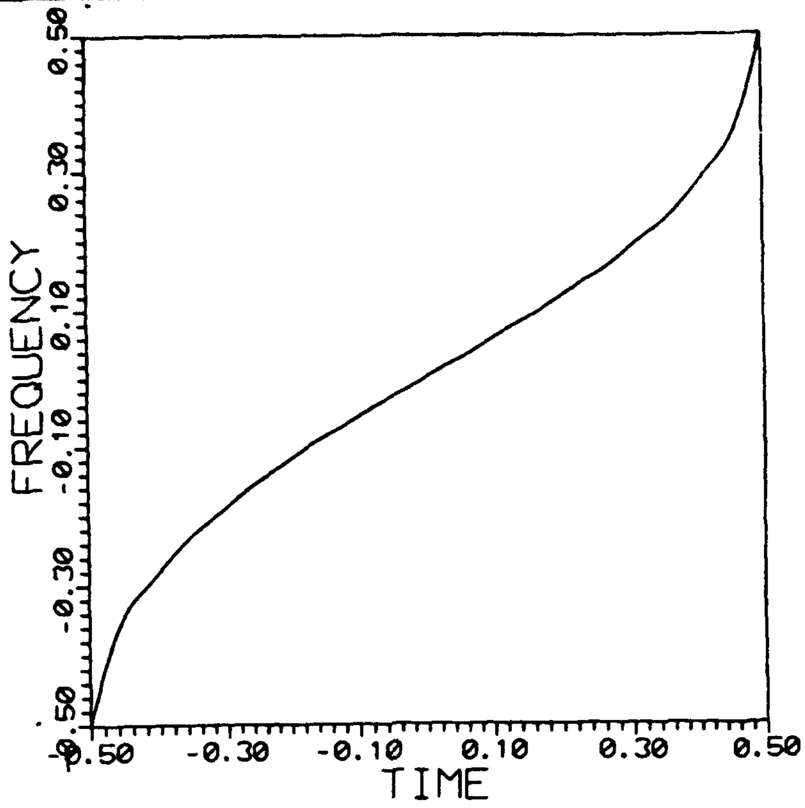

Fig. 3a The frequency function of a nonlinear FM for a 40 dB Taylor weight

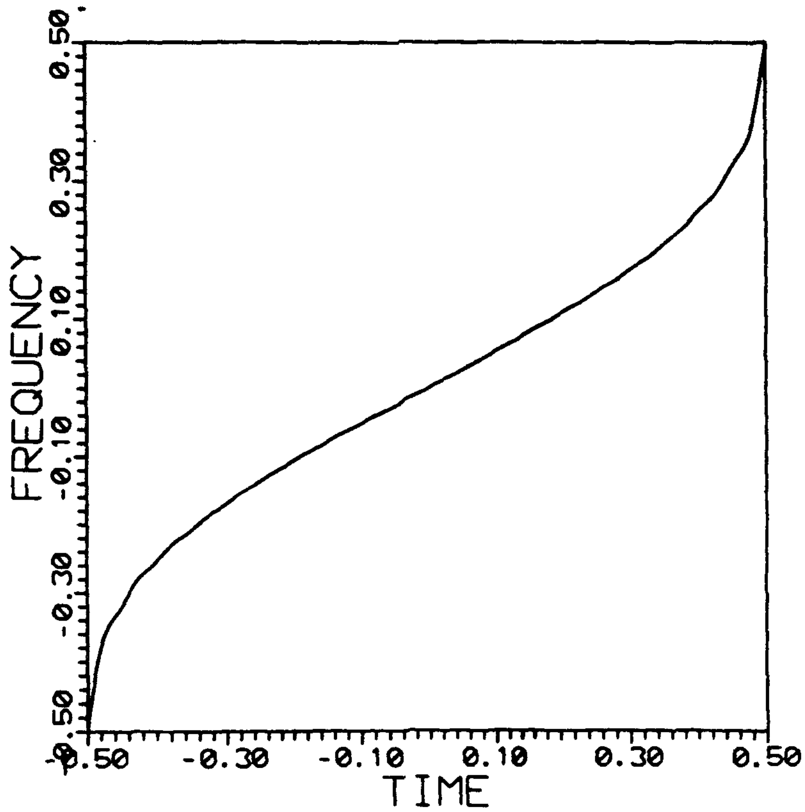

Fig. 3b The frequency function of a nonlinear FM for a 50 dB Taylor weight

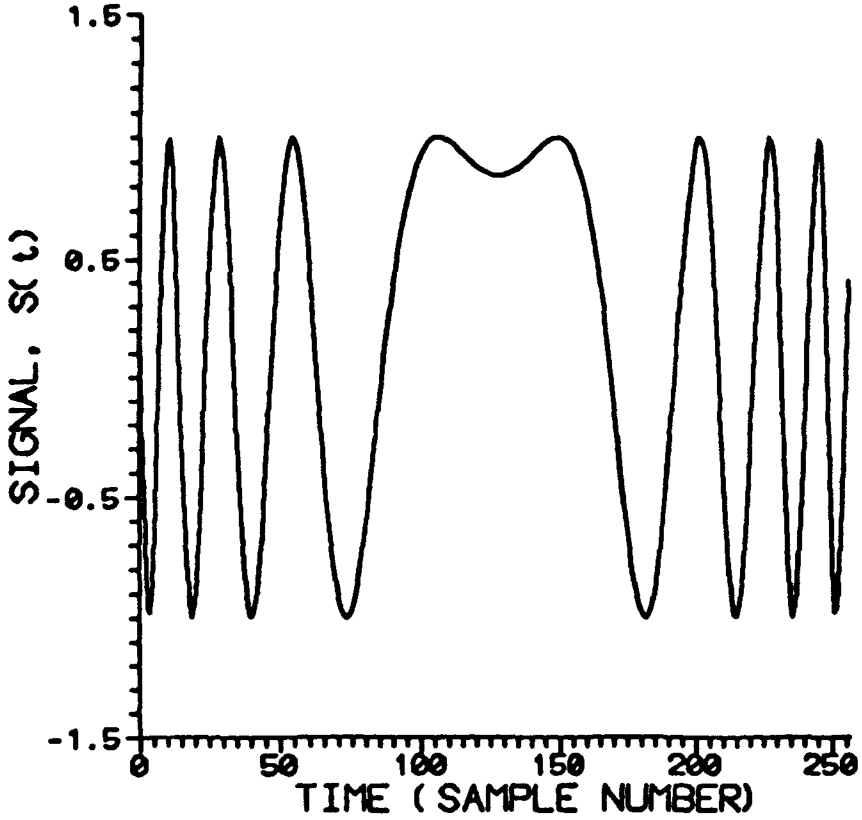

Fig. 4a The nonlinear FM signal for $\rho=50$

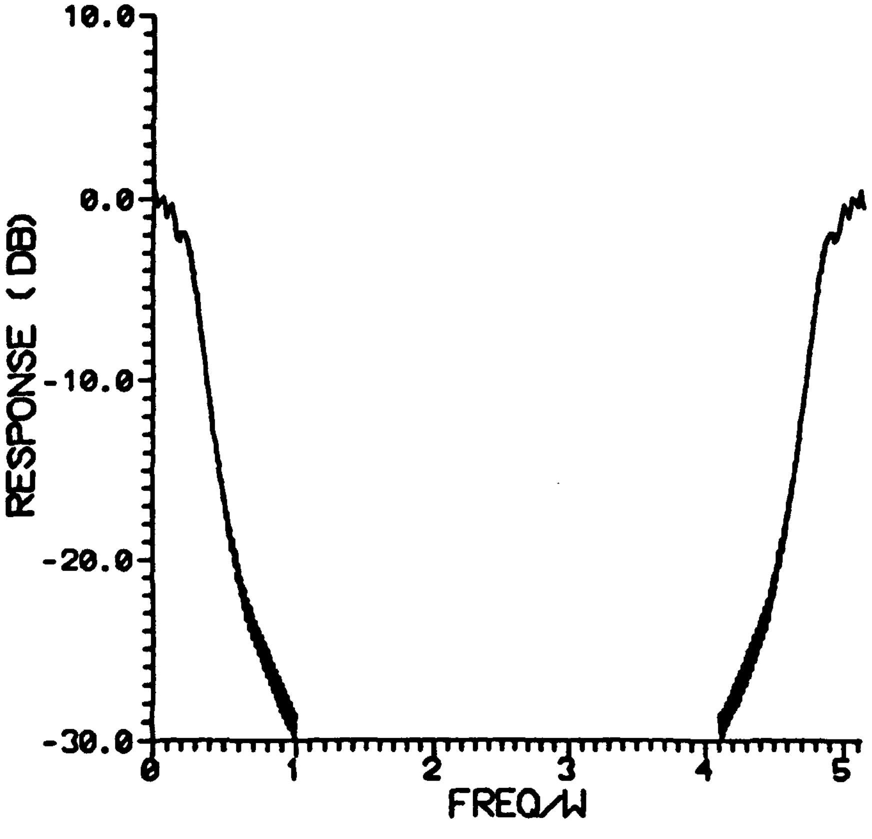

Fig. 4b The nonlinear FM signal spectrum for $\rho=50$

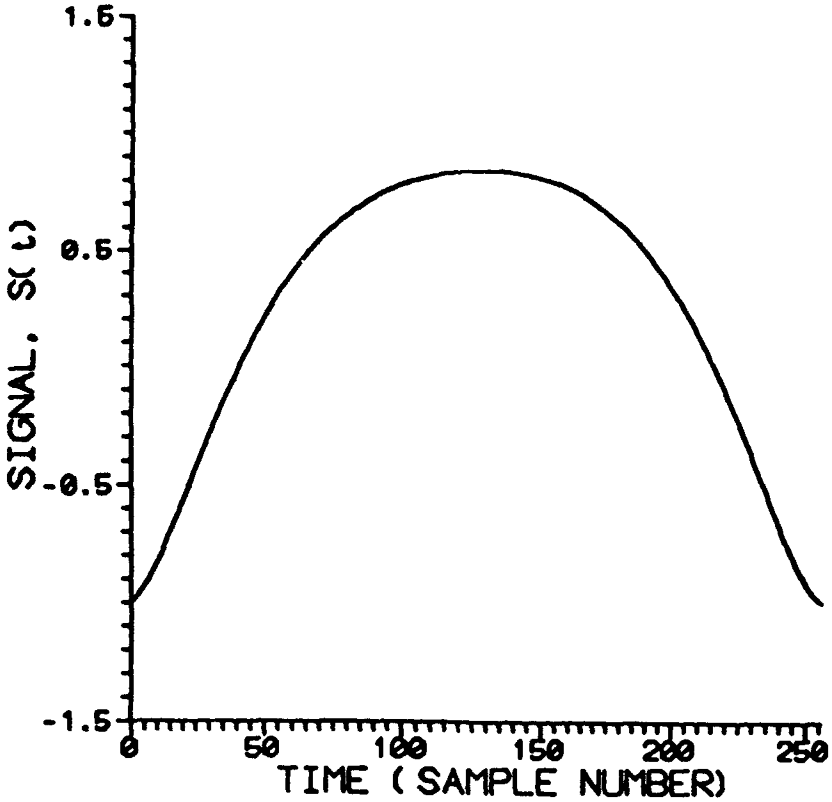

Fig. 5a The nonlinear FM signal for $\rho=5$

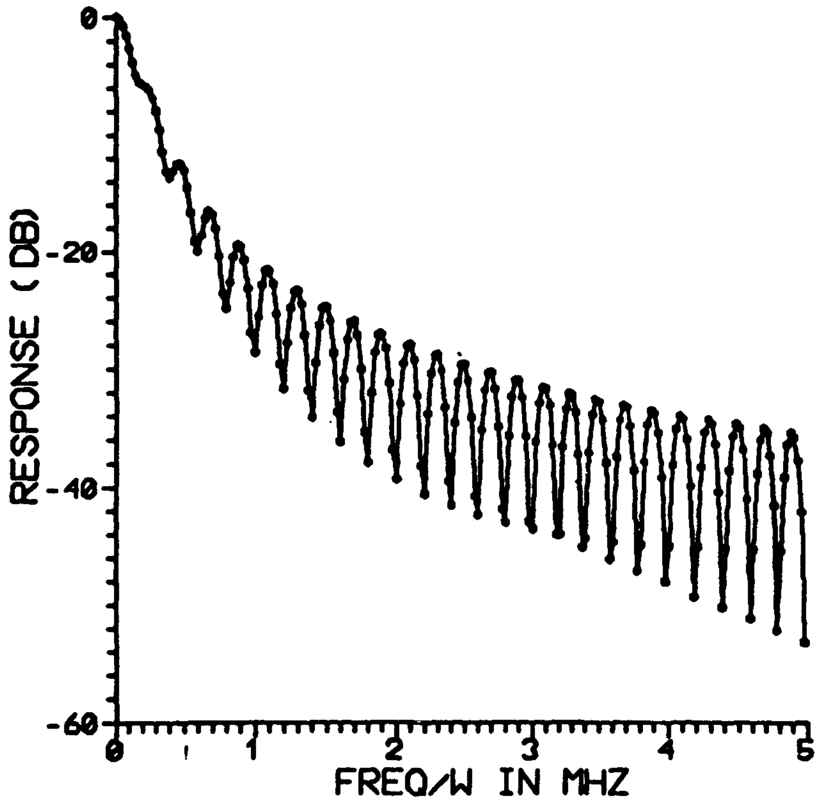

Fig. 5b The nonlinear FM signal spectrum for $\rho=5$

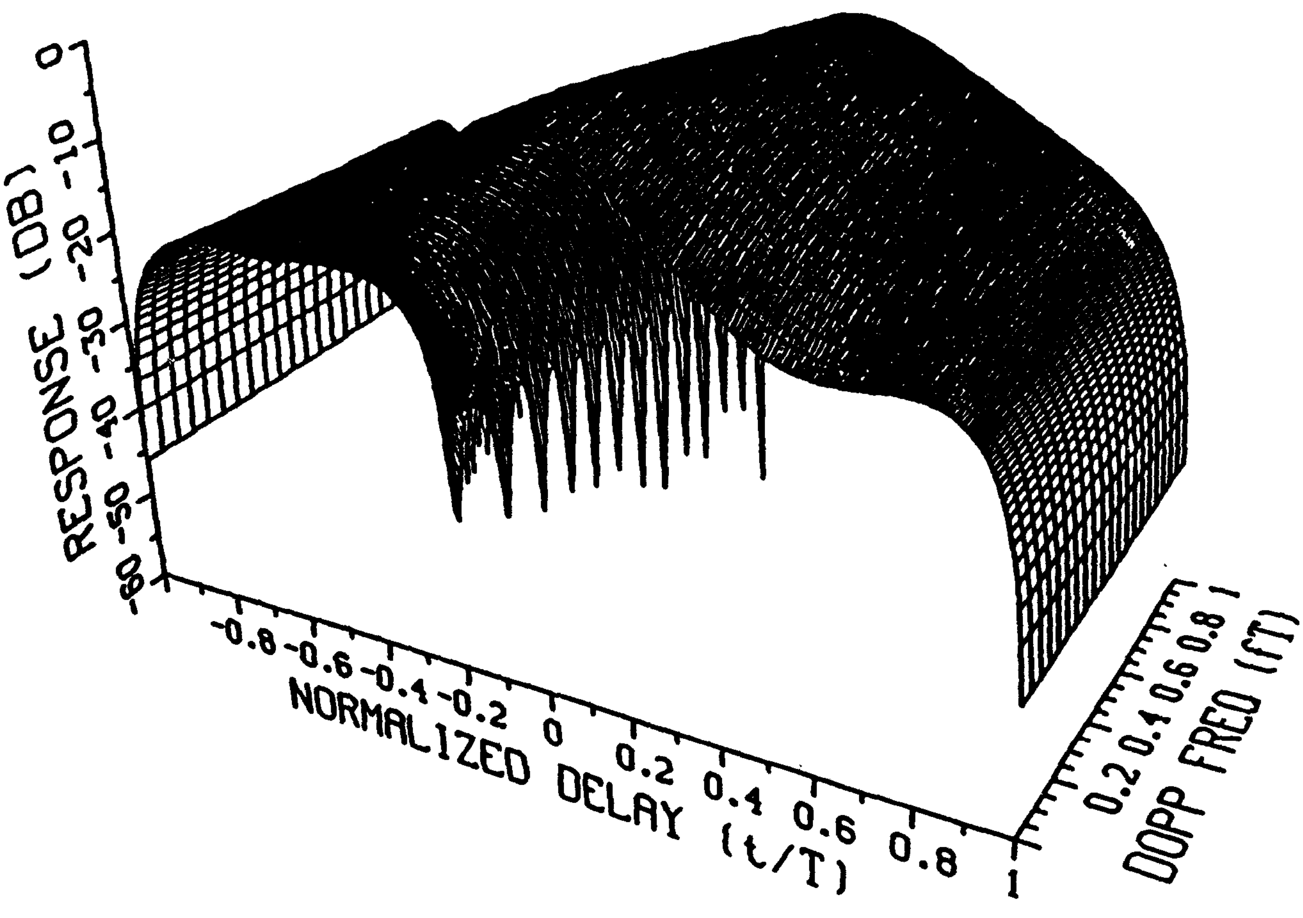

Fig. 6 The ambiguity diagram of a nonlinear FM signal for $\rho=5$

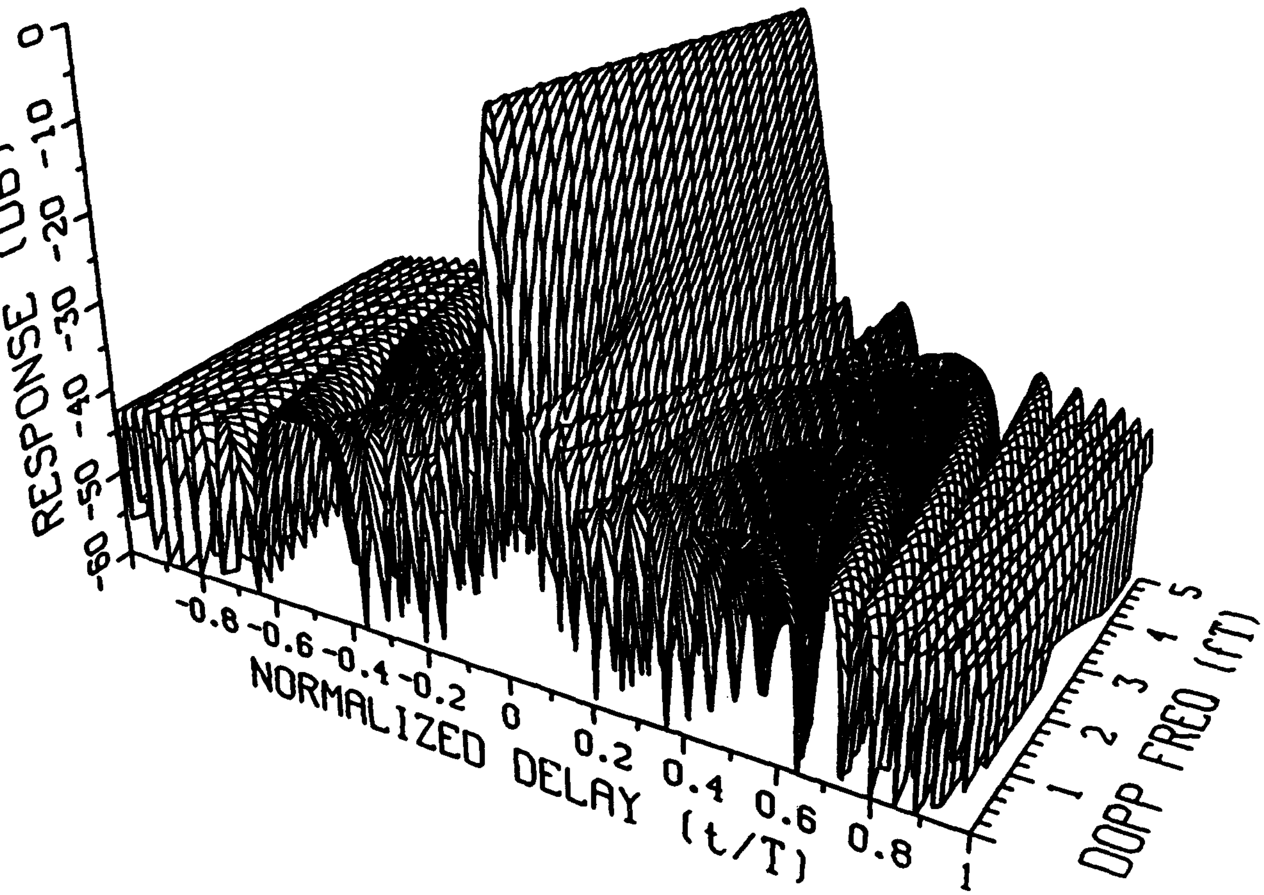

Fig. 7 The ambiguity diagram of a nonlinear FM signal for $\rho=50$

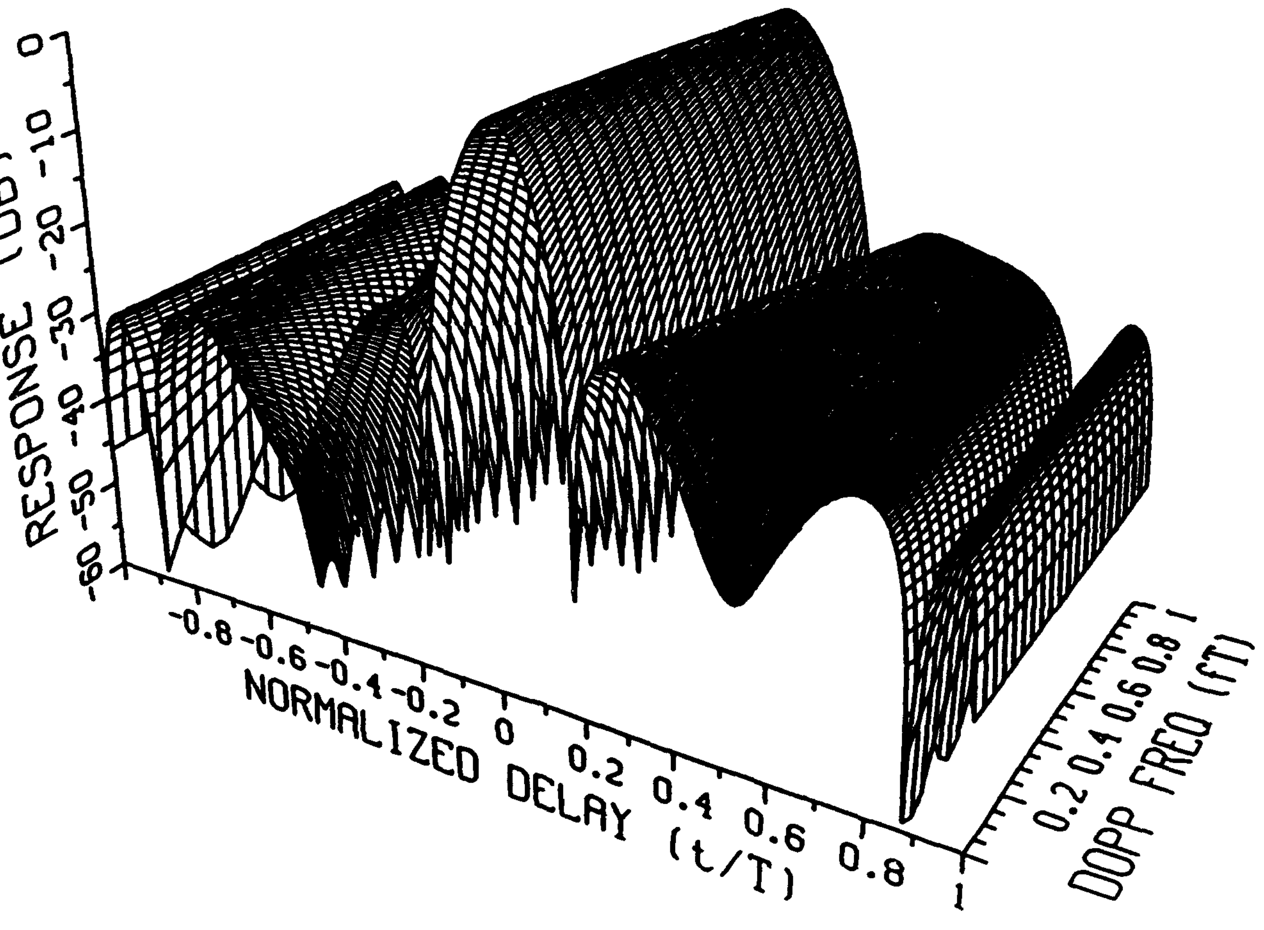

Fig. 8 The ambiguity diagram of a nonlinear FM signal for $\rho=13$

appropriate sampling rate will be fed into the "Mitchell filter" followed by a matched filter to extract the compressed signal.

We consider that the nonlinear FM has a compression ratio of 5 and a bandwidth of 1 MHz throughout the report. For simplicity, we assume at the beginning that the transmitted signal has a carrier frequency of 30 MHz. The signal is generated from Eq. (4) with $\hat{f}_o$=30 and ρ=5, and is composed of 3059 samples. Fig. 9 shows the spectrum of this signal, where the carrier is clearly located at 30 MHz. Apparently the nonlinear FM spectrum characteristics are preserved (Fig. 9b). Note that, with the application of FFT in obtaining Fig. 9, the frequency extends to $(1/\Delta\hat{t})\times(w/\rho)$ which is 611.6 MHz for ρ=5, W=1 MHz and $\Delta\hat{t}$ =1/3058. To have appropriate resolution in performing "direct-conversion-to-baseband" at a later time, the signal samples are padded with 29,709 zeros to have a total of 2^{15} points. Therefore the frequency resolution shown in Fig. 9 is $611.6/2^{15}$, or 0.0187 MHz.

The above signal is transmitted through a bandpass filter of 10 MHz centered at 30 MHz. Here we use a Butterworth band-pass filter $H(\hat{s})$ of order 8. The filter is obtained from a normalized low-pass Butterworth filter H(s) through a transformation of $s=(\hat{s}^2+\omega_o^2)/B\hat{s}$ where ω_o is the filter center frequency and B is the desired bandwidth. We define H(s)=1/D(s) and D(s) is the Butterworth polynomial such that,

$$D(s) = (1+0.3902s+s^2)\cdot(1+1.1111s+s^2)\cdot(1+1.6639s+s^2)\cdot(1+1.9616s+s^2)\,. \tag{5}$$

Figure 10 shows the frequency responses of the Butterworth filters at (1) ω_o =30 MHz and B=10 MHz, (2) ω_o =30 MHz and B=2 MHz and (3) ω_o =2.5 MHz and B=4 MHz. Figure 11 is the transmitted nonlinear FM signal after passing through the 1st filter described above. In the receiver, the signal is first band-limited by a 2 MHz Butterworth filter centered at 30 MHz and down-converted to a low IF at 2.5 MHz followed by a 4 MHz Butterworth filter. The resulting signals appearing at the output of these two filters are shown in Figs. 12 and 13, respectively. Note, at this stage, ripples are observed near and beyond the instantaneous bandwidth of 1 MHz.

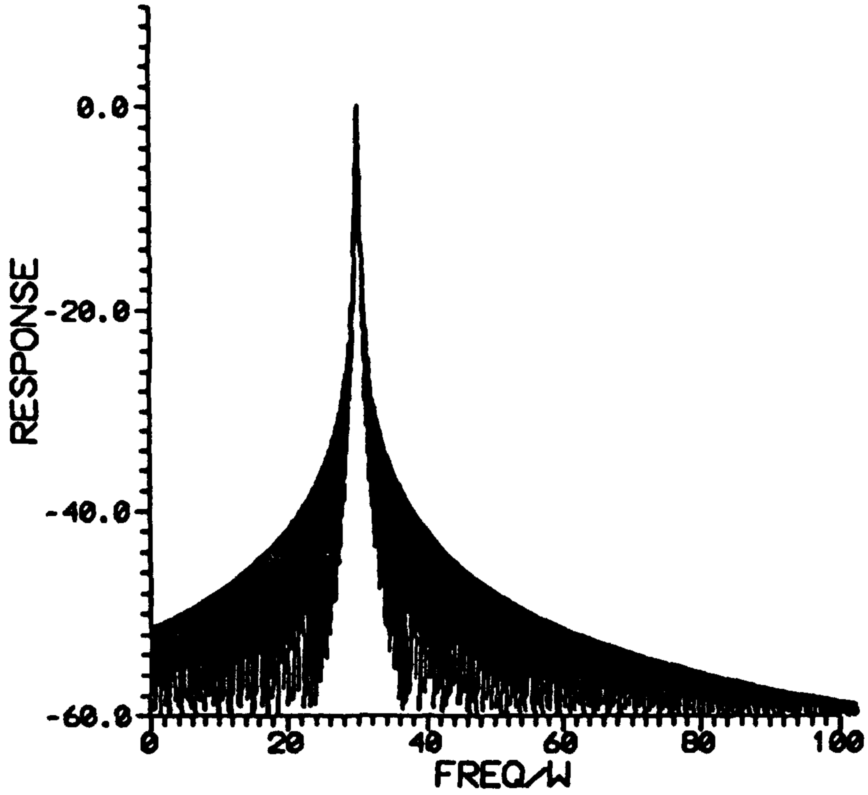

Fig. 9a The nonlinear FM signal spectrum with a carrier of 30 MHz

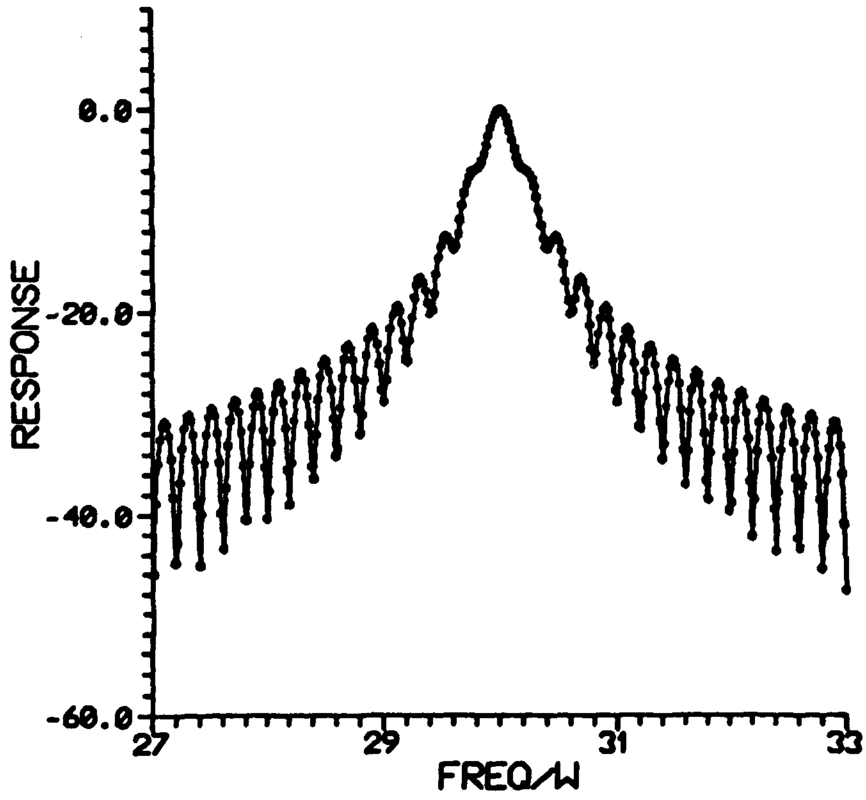

Fig. 9b The nonlinear FM signal spectrum with a carrier of 30 MHz (Fine Scale)

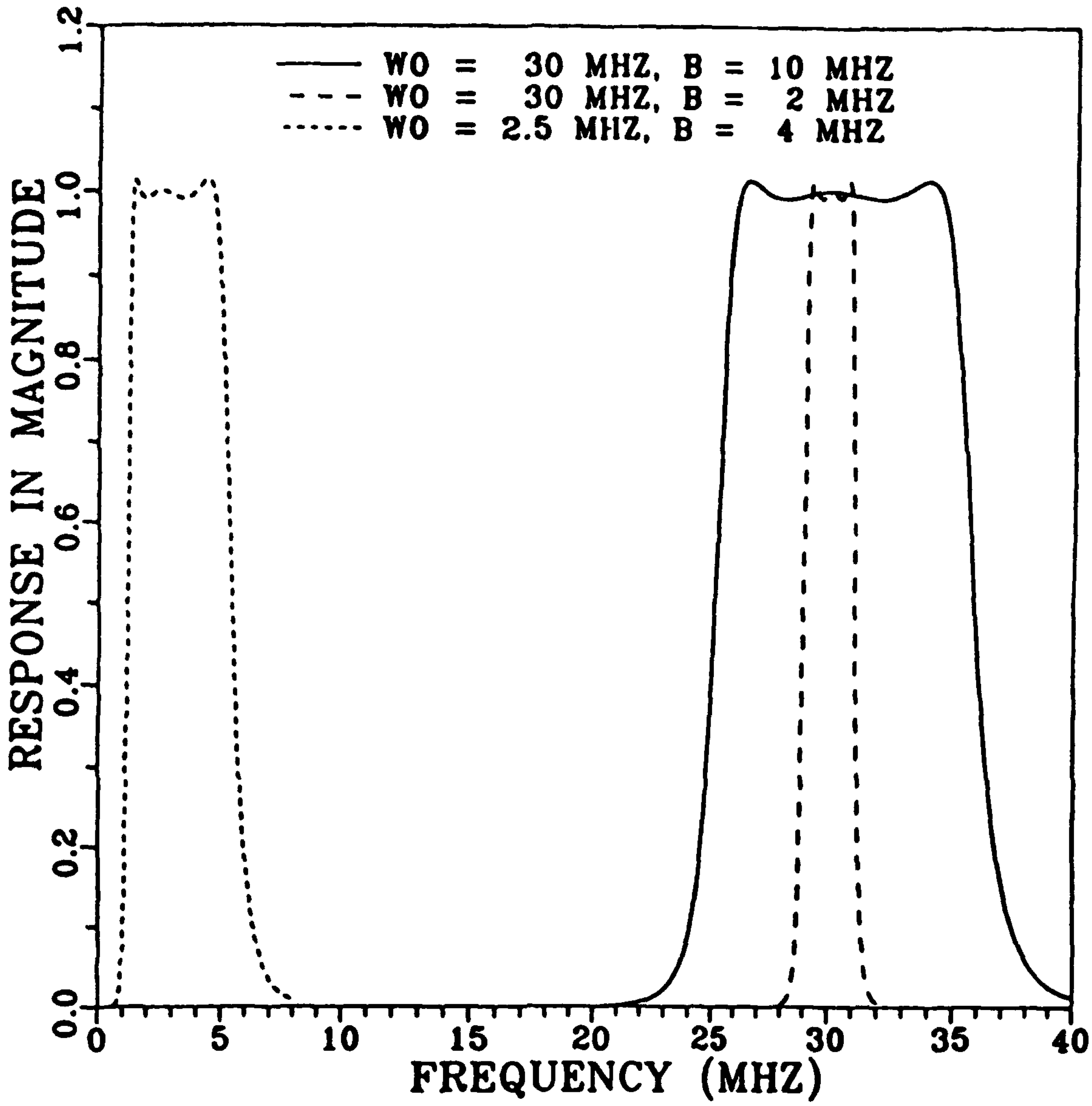

Fig. 10 The Butterworth filter responses for n=8

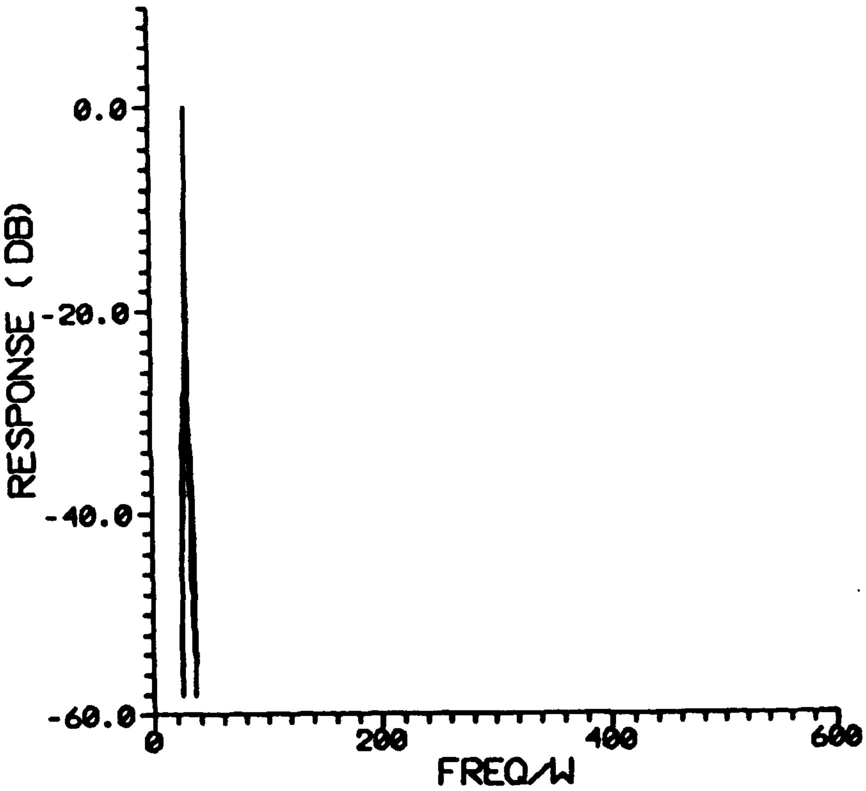

Fig. 11 The spectrum of transmitted signal

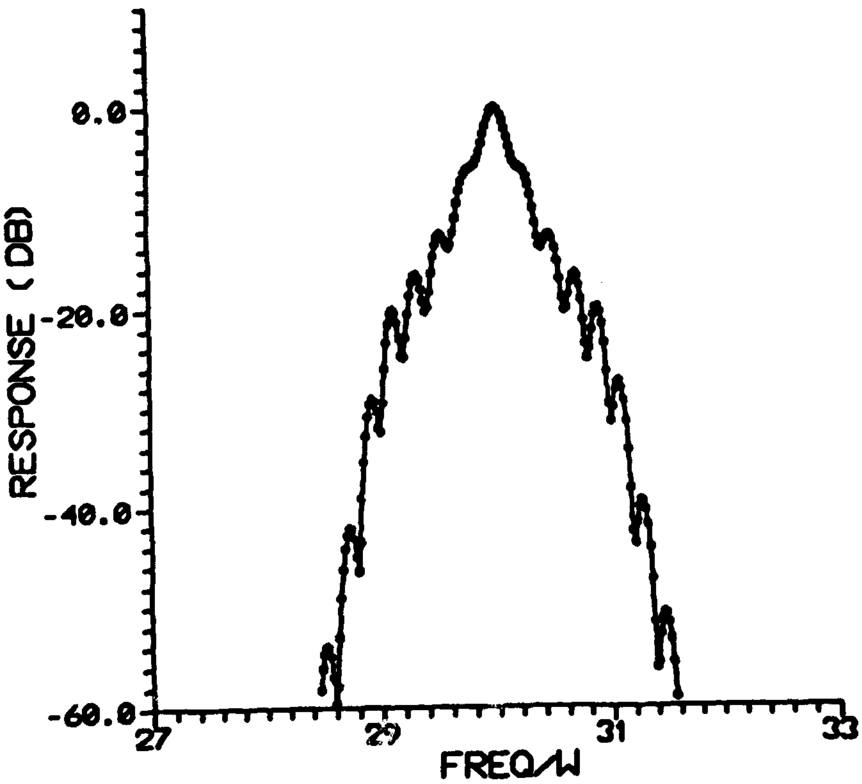

Fig. 12 The spectrum of received nonlinear FM signal (after the 2 MHz Butterworth filter)

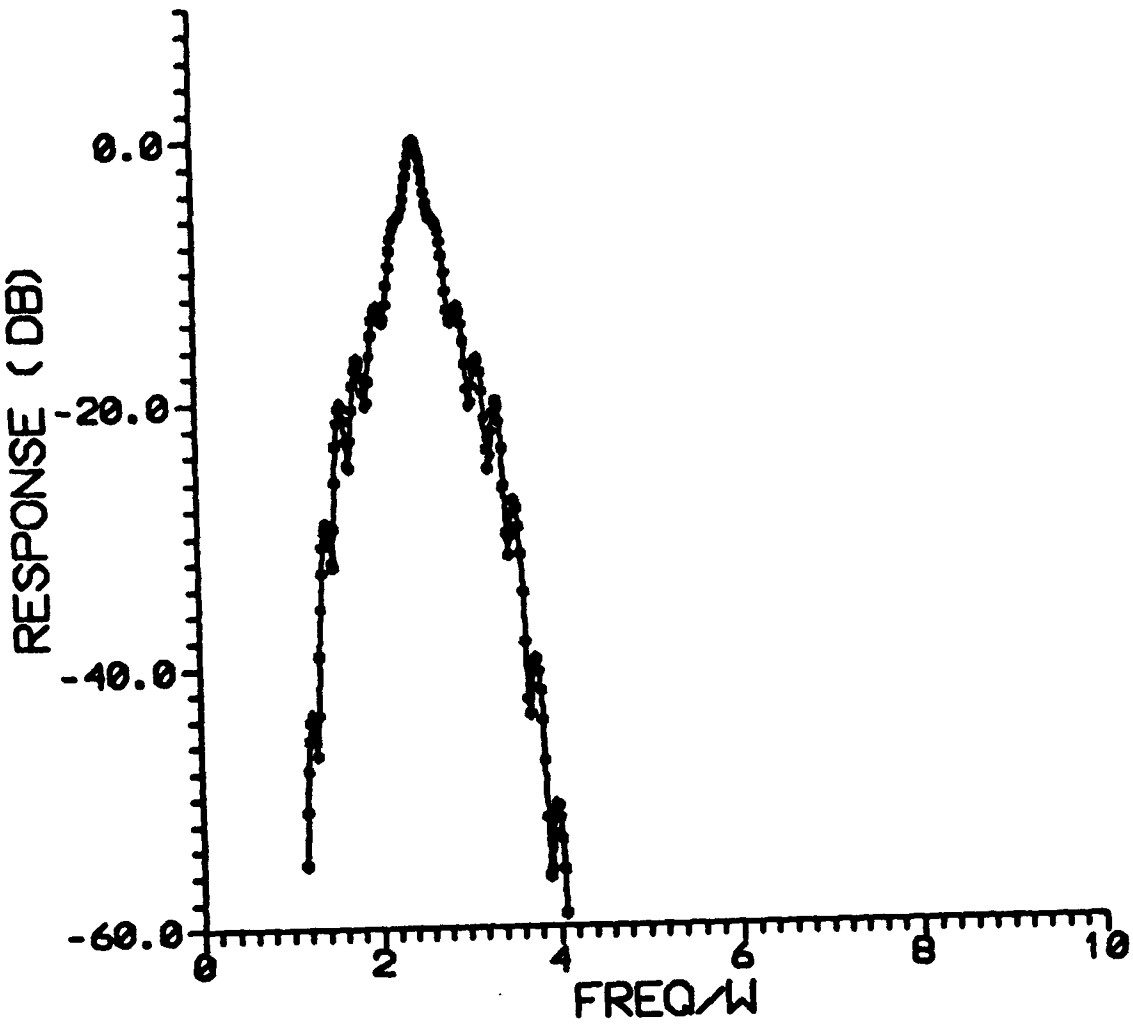

Fig. 13 The spectrum of received nonlinear FM signal (after the mixer shown in Fig. 2)

4. DIRECT CONVERSION TO THE BASEBAND

It is often difficult to match the gains, phases and frequency responses of two processing channels which produce the in-phase and quadrature components of a received band-limited RF signal. A simple way of generating these complex components is a direct conversion of the signal to the baseband using a combination of mixing, sampling and filtering. As discussed in Ref. 4, the RF signal needs to be converted to a very low IF with frequency near the bandwidth, and then sampled and digitized at 4 times this frequency before processing through a digital filter.

In the previous Section, the received nonlinear FM has been deliberately mixed with a reference signal of 27.5 MHz and down-converted to a low IF of 2.5 MHz. The IF signal is then sampled at a rate of 4×2.5, or 10 MHz. The resultant spectrum is shown in Fig. 14 where the signal at negative frequencies repeats at the location of 7.5 MHz. We recall that the frequency resolution is 0.0187 MHz when the spectra of Figs. 11 through 14 are interpreted. Consequently there are about 536, i.e., 10/0.0187, samples shown in Fig. 14. Since only real-value samples are available after the sampling and digitizing processes, the A/D outputs are actually the real part of the inverse FFT of the spectrum of Fig. 14. Figure 15 shows these time-series samples. In the figure, the normalized sampling period $\Delta t/T$ is 1/(10x5) (obtained by substituting $\rho=5$, $f=10$ MHz and $W=1$ MHz in the expression $W/f\rho$), and the normalized time has extended to 10.7, i.e., 1/(0.01867x5) or 0.02X536. Apparently there are only 536/10.7 or 50 meaningful samples.

Based on the principle of a Hilbert transform, an infinite duration impulse response (IIR) filter can be used to create the baseband complex signals [3,4]. As practical applications, a simpler FIR filter, instead of an IIR, is suggested in [5]. This digital FIR filter is described by

$$y_k = x_k - j\left[c_1(x_{k+1}-x_{k-1}) + c_3(x_{k+3}-x_{k-3}) + \dots + c_{2n-1}(x_{k+(2n-1)}-x_{k-(2n-1)})\right] \qquad (6)$$

where x_k is the real sample input to the filter at discrete instant k, and y_k is the complex output of the filter. It is shown in [5] that, for n=4, $c_1=0.610741$, $c_3=0.144644$, $c_5=0.041140$ and $c_7=0.007475$.

To further obtain the in-phase and the quadrature components of the baseband signal, we compute every fourth sample of the above

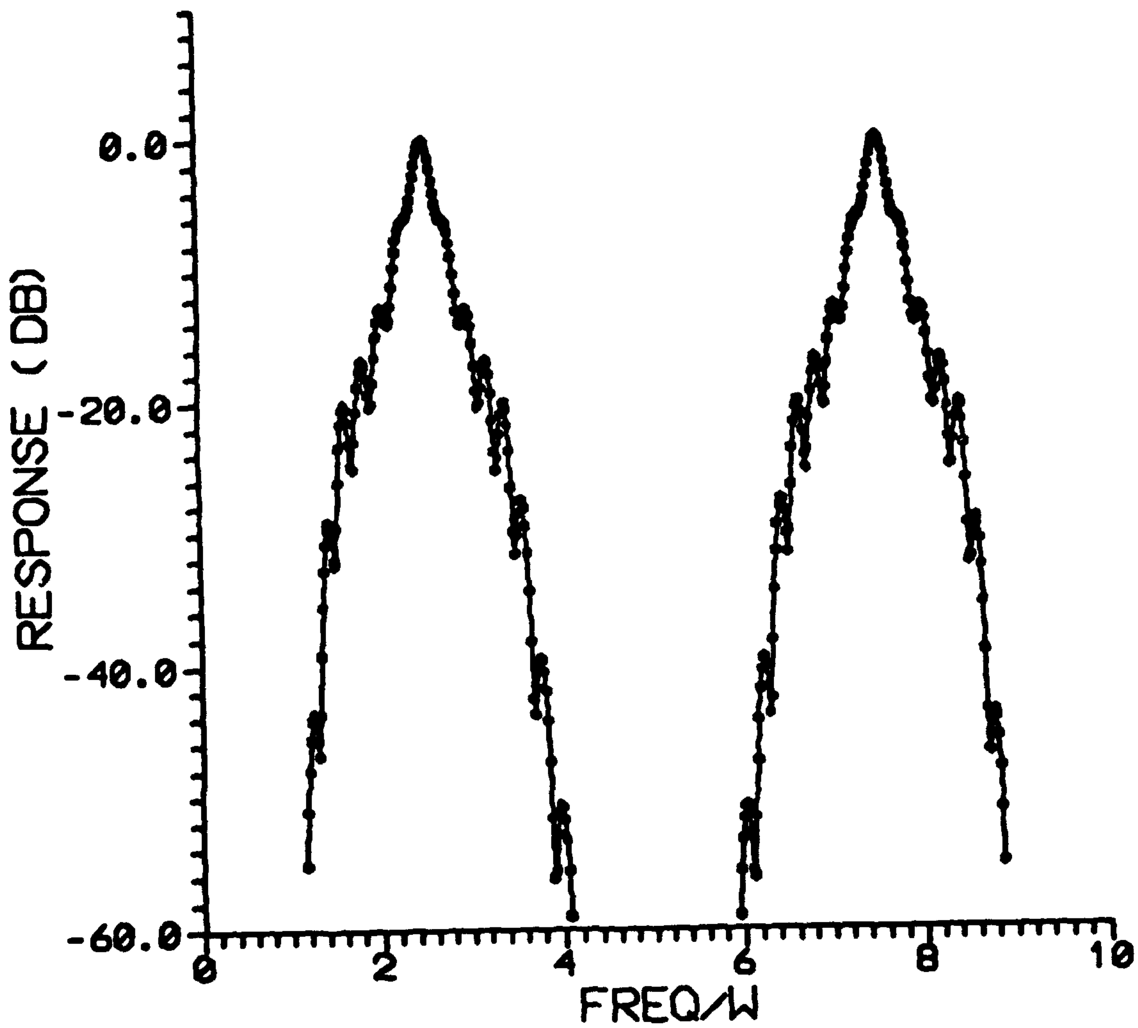

Fig. 14 The spectrum of received nonlinear FM signal (after down-converted, filtered and sampled)

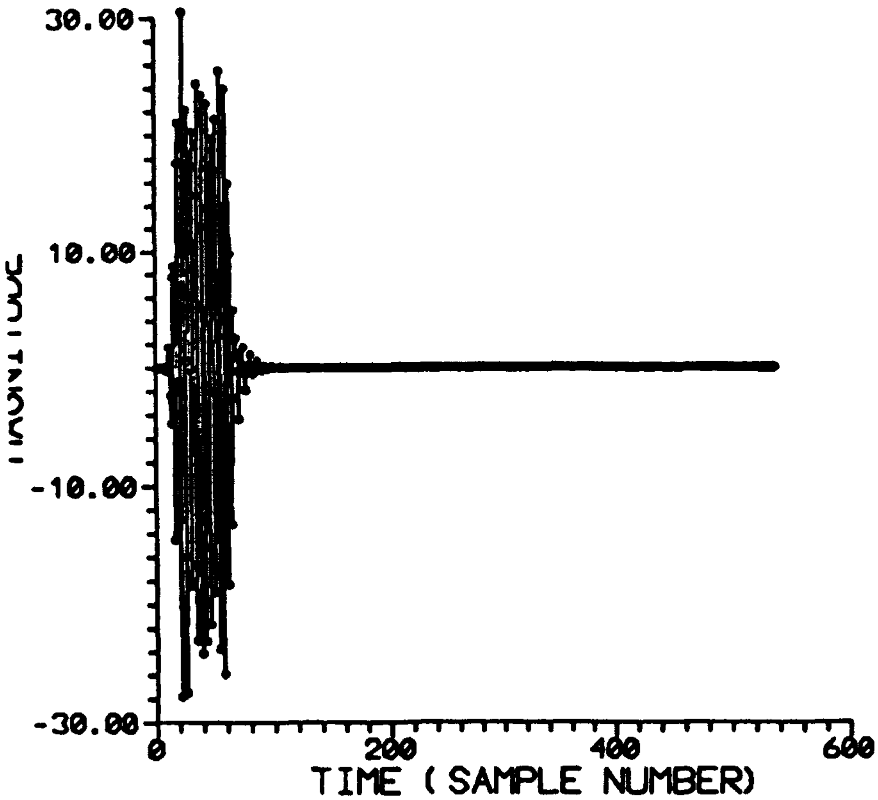

Fig. 15a The received sampling signal

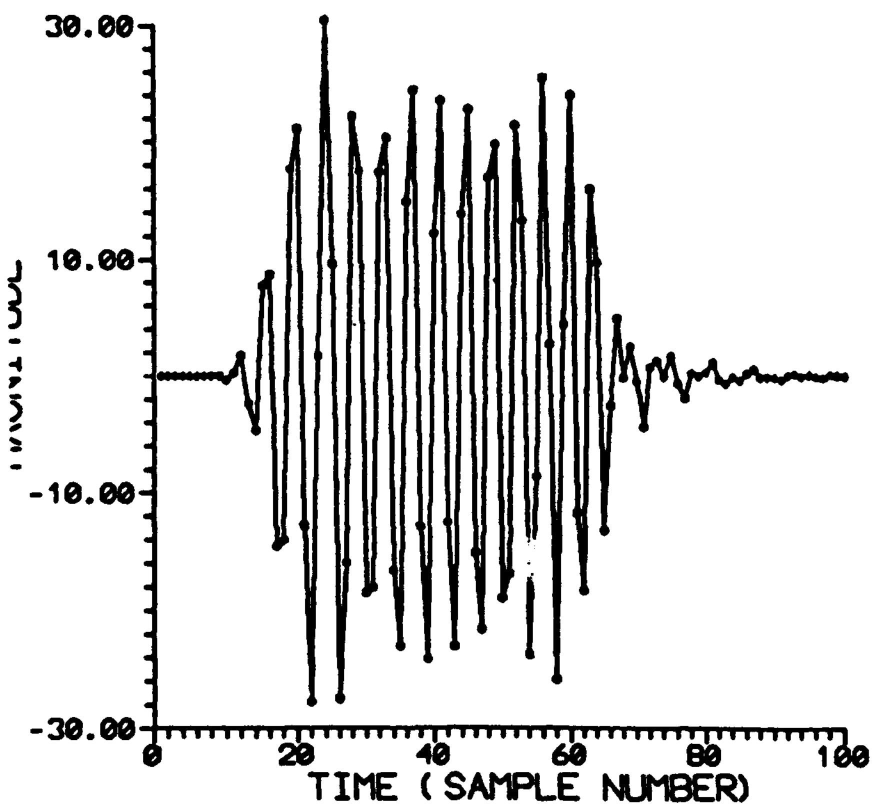

Fig. 15b The received sampling signal (Fine Scale)

filtered sequence $\{y_k\}$, or decimate the filter sequence by a factor of 4. The results are shown in Figs. 16 and 17 (for the first 30 points). There are actually $(536-2\times7)/4$, or 130, reconstructed signal samples available in both figures due to decimation. Among them, about 12 $(130\div10.7)$ samples appear useful. The deduction of 2×7 samples (for $n=4$) is due to the extra points needed for the filter forward/backward prediction in the beginning and at the end of the process. It can be seen that the pattern of Fig. 16 is quite similar to that of Fig. 5a. The shapes of both Figs. 16 and 17 vary, depending on the selection of initial point during the decimation process. However, these patterns in pair preserve the nonlinear FM characteristics.

5. MATCHED FILTER OUTPUT

Using the "direct-conversion-to-baseband" filter (or the "Mitchell filter"), we generate the complex components of a nonlinear FM signal. These reconstructed signal samples are then passed through a compression network so that a compressed signal is retrieved. Here we only consider the case that the receiver (or the matched filter response) is matched to the reconstructed signal. Figure 18 shows the nonlinear FM spectrum for the reconstructed signals obtained in Figs. 16 and 17. In the figure, the x-axis in frequency has a range from 0 to $(1/\Delta\hat{t})\times(W/\rho)$, where $\Delta\hat{t}=\Delta t\times4=0.08$, $\rho=5$ and W indicates the bandwidth. Figure 19 plots the resulting matched-filter output in which the range time sidelobe is approximately -18 dB. By comparing Figs. 16, 18 and 19 with Figs. 5a, 5b and 6, respectively, we observe that the nonlinear FM signals are well reconstructed.

6. DISCUSSIONS AND SUMMARY

In this report, we examine the nonlinear FM signal and the direct-conversion-to-baseband" through a simple process of mixing, sampling, and digital filtering. In application to a radar system of interest, a band-limited signal of low pulse compression ratio is used to demonstrate the reconstruction of the signal at baseband and also to show the matched-filter output results. We generate a phase predistortion function which modifies the transmitted linear FM signal and synthesizes the desirable effect of a sidelobe weighting function. The predistortion phase characteristics assure that the envelope of the transmitted signal is rectangular and the

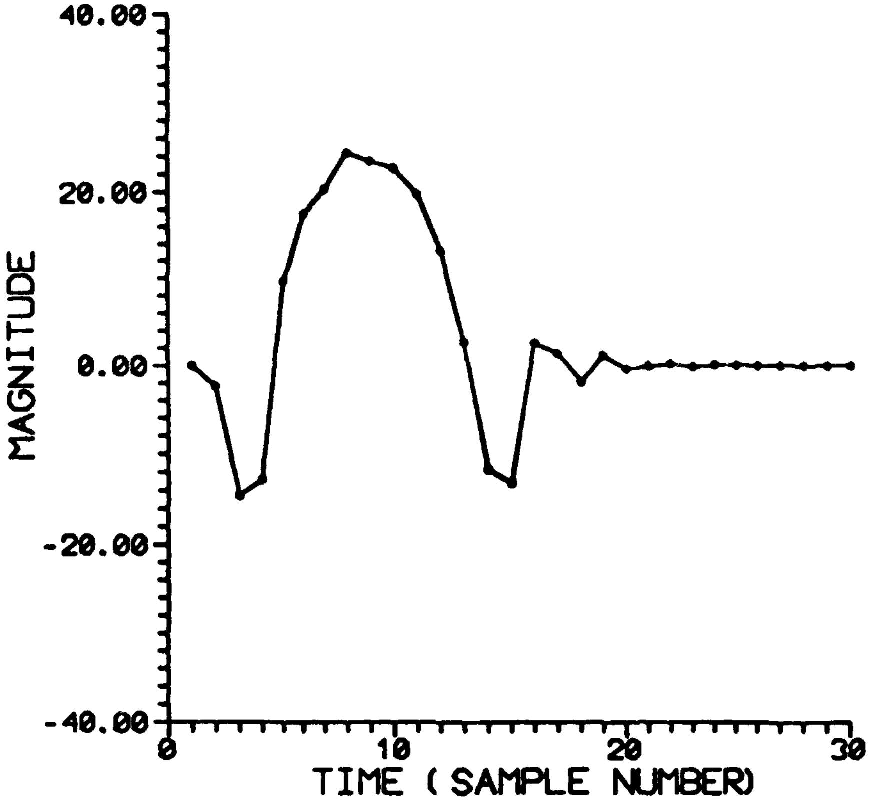

Fig. 16 The in-phase component of the reconstructed signal

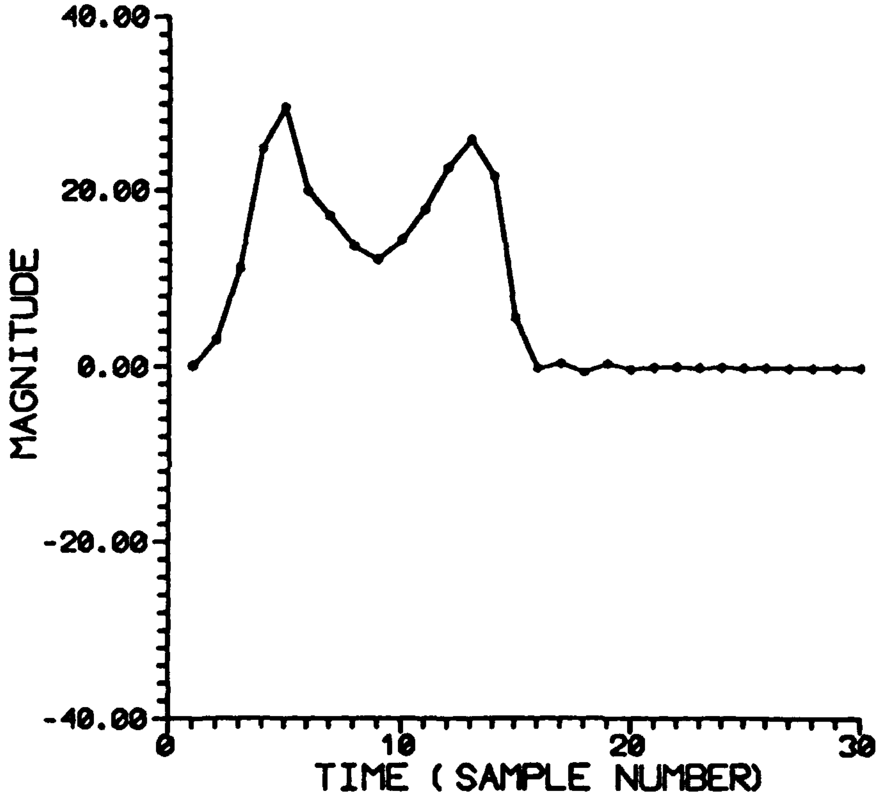

Fig. 17 The quadrature component of the reconstructed signal

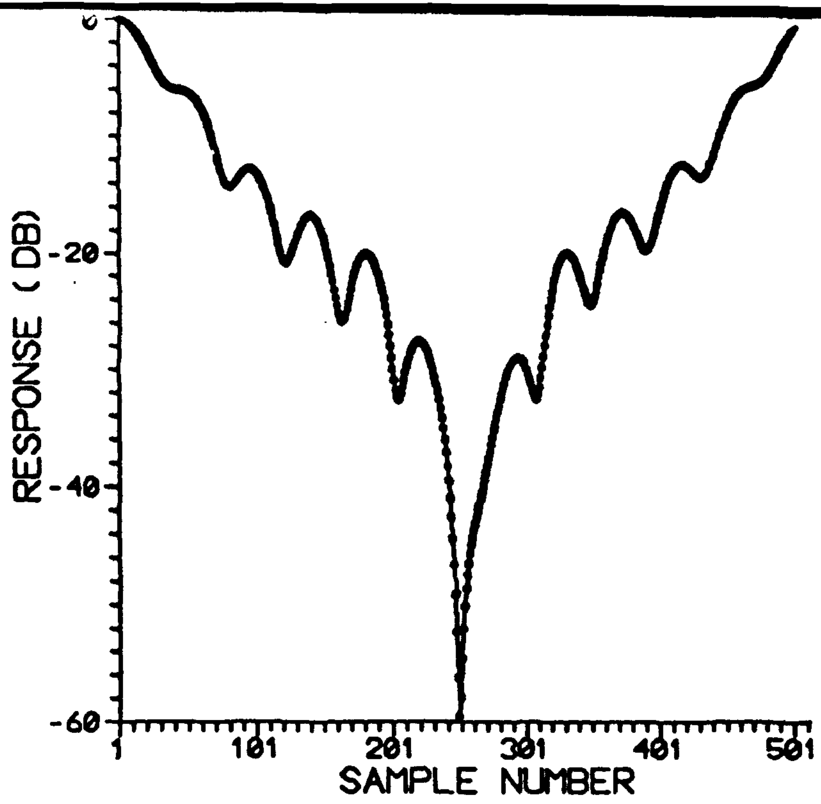

Fig. 18 The spectrum of reconstructed signal (obtained from Figs. 13 and 14)

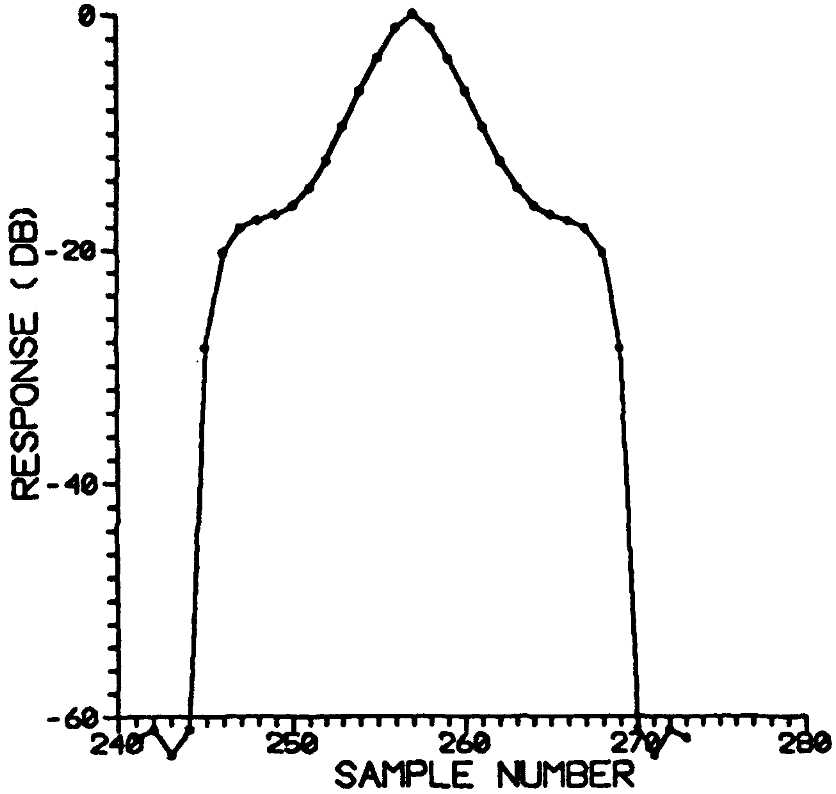

Fig. 19 The matched-filter output

spectrum is Taylor shaped. A generalized nonlinear FM signal is derived and listed in Table 1 for various Taylor weightings.

To gain insight into the reconstruction of a complex signal through the "direct-conversion-to-baseband", we simulate the process using a 30 MHz transmitted nonlinear FM. The signal is first received and bandpassed by a 2 MHz Butterworth filter, then down-converted to a low IF of 2.5 MHz. To avoid a mismatch in obtaining the signal in-phase and quadrature components from two separate processing channels, we directly convert the low IF to its baseband by a combination of sampling/digitizing, filtering and decimation. The compressed signal is also obtained through a pulse compression filter matched to the transmitted signal.

For a radar using a nonlinear FM signal, we practically consider the waveform having a prf of 5 or 10 KHz. Assume that the duty cycle and the instantaneous bandwidth of the signal are held constant at 5% and 625 KHz, respectively. Then the uncompressed pulsewidth of 10 μs or 5 μs has to be compressed to a duration of 1.6 μs. The resultant pulse compression ratio, or the time-bandwidth product, is 6.25 or 3.125. If the maximum Doppler is about 10 KHz, the maximum Doppler shift normalized to the bandwidth of 625 KHz is only 0.016. It appears that the matched-filter output may not be affected very much by the Doppler shift up to 0.016. However, the small pulse compression ratios as compared to ρ=50 or greater significantly limit the reduction of range time sidelobe levels.

We demonstrate here that a nonlinear FM signal can be directly converted to the baseband through a combination of mixing, sampling and filtering processes for the case ρ=5 and B=1 MHz. A band-limited signal shows little effect on the retrieval of the compressed signal. The range time sidelobe of approximately -18 dB is achieved for the above case. It is recommended that a weighting function following the compression network should still be used for the nonlinear FM case. To achieve further reduction on sidelobe levels, other alternatives may include the synthesis of the compression network (incorporated with a hardware design) which would result in a favorable combining transfer function from $S_I(\omega)$ to $S_U(\omega)$ in Fig. 1.

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